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# An Eight-Switch Five-Level Current Source Inverter

WeiQi Wang, *Student Member, IEEE*, Feng Gao, *Senior Member, IEEE*, Yongheng Yang, *Senior Member, IEEE*, and Frede Blaabjerg, *Fellow, IEEE*

**Abstract** — This paper proposes an eight-switch three-phase five-level current source inverter (CSI), which employs only one traditional H6 inverter and two shunt branches at the DC side to realize the five-level switching. The corresponding space vector modulation (SVM) strategy for the proposed CSI topology is also presented, which uses the two shunt-connected power switches to add certain special modulation-state segments to ensure the switching instants completed under lower current stresses and lower power dissipation. Compared with the state-of-the-art CSI solutions, the proposed topology has a comparable hardware cost as the three-level H6 CSI, while outputs five-level currents. The low output THD may help to reduce the sizes of the passive components in the system, and the modulation scheme reduces the switching and conduction losses of the semiconductor switching devices in H6 CSI module, which can make it possible to increase the output current rating of the system in a certain degree. Simulation and experimental results verified the performance of the proposed CSI.

**Index Terms** — current source inverter; multi-level converter; modulation.

## I. INTRODUCTION

In general, inverter topologies can be categorized into two types—voltage source inverter (VSI) and current source inverter (CSI). In the past and until now, VSI has been the dominant inverter topology. At the same time, attempts toward advancing the CSI have never been stopped, and the penetration of CSI topologies continuously challenges the VSI market, mainly due to the uniqueness of CSI topologies (e.g., the inherent short-circuit protection ability, high voltage boost capability, and superior  $dv/dt$  performance). On the contrary, these unique aspects are the critical issues for the VSI applications [1], [2]. In this respect, the CSI topologies are

promising, especially in certain applications, e.g. high power electric drives [3] and photovoltaic (PV) systems [4], requiring a high voltage boosting capability.

Additionally, to reduce the dependence of bulky output filters for less harmonic emissions and also to lower the voltage stress on power devices, multilevel technologies have firstly been introduced to VSI systems. As such, low-rating power devices can be used (contributing to cost reduction), and also potentially, a high efficiency can be achieved. Thus, many multilevel VSI systems have been commercialized in the past few years, among which the neutral point clamped (NPC) multilevel inverters, flying capacitor (FC) multilevel inverters, cascaded H-bridge (CHB) multilevel inverters, and modular multilevel converters (MMC) are the favorites [5]–[7]. In a similar way, the multilevel switching characteristics can be a great added value to CSI topologies [8]. In recent years, many attempts have been made to improve the multilevel CSI (MCSI) technologies. For instance, the single-rating inductor MCSI was proposed by employing multiple H6 CSI modules connected in parallel on the AC side [9], [10]. However, current circulating and imbalance issues are associated with this kind of MCSI topology. Furthermore, to remove the passive elements of the modules, the multi-rating inductor MCSI was introduced in [11] to alleviate the current circulating and imbalance problem and a paralleled-MCSI with independent DC-links was introduced in [12]. The paralleled-MCSI topology effectively addresses the issues of circulating currents and unbalances, whose corresponding modulation schemes were proposed in [13] and [14]. Moreover, an improved paralleled MCSI topology using a shared DC-link was presented in [15]. In addition, certain variants like the buck-boost derived MCSI [16] have been developed to extend the operating range. To enable the high-power operation, a current source modular multilevel converter (CS-MMC) was proposed in [17], which utilizes multiple current source cells connected in parallel with a significant increased number of power devices and inductors.

With the above considerations, this paper first reviews the general configuration characteristics of various MCSI topology solutions. Then, this paper proposes an eight-switch three-phase five-level CSI topology with the corresponding space vector modulation (SVM) strategy, which enables the eight-switch CSI topology output the five-level switching current but using fewer power switches, thereby lowers hardware cost. It should be pointed that the SVM scheme is much simpler than the conventional methods, being another advantage of the proposed inverter. The rest of this paper is organized as follows. In Section II, the conventional MCSI topological attempts are briefly introduced. In Section III, the circuit configurations for the proposed eight-switch three-phase five-level CSI topology are proposed, and then the compatible modulation strategies using the SVM are presented

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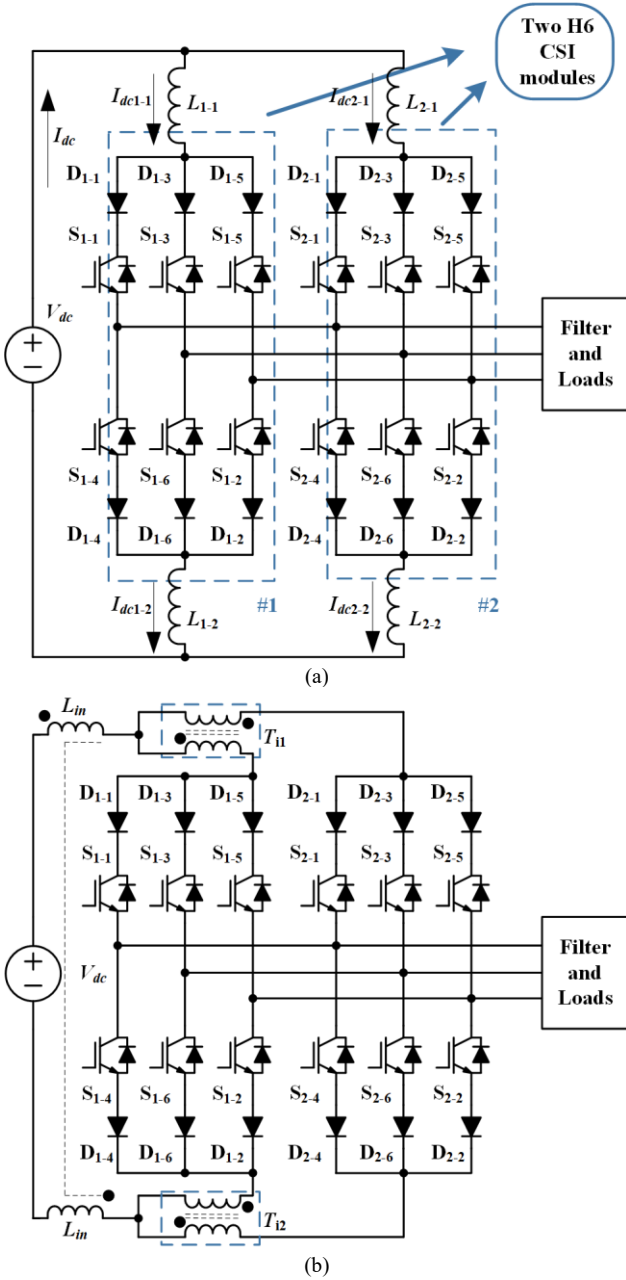


Fig. 1. Three-phase single-rating inductor MCSI topologies: (a) with four inductors of the same rating and (b) with two interleaved inductors (coupled inductors).

in detail. The modulation sequences as well as the characteristics of operation are also analyzed. Finally, Matlab simulation and the experimental tests are presented to verify the performance of the proposed topology.

## II. TOPOLOGIES AND MODULATION OF THREE-PHASE MULTILEVEL CSIS

A conventional CSI consisting of six power switches is the simplest and most fundamental CSI. The DC current flows through two power switches, and commutates between lateral arms, which has six active states and three null states. To reduce the current stress on each power switch, multi-level technologies were introduced to CSI. There are several types of three-phase MCSI topologies, i.e., the single-rating inductor MCSI, the multi-rating inductor MCSI, the paralleled H-bridge MCSI, the buck-boost MCSI, and the CS-MMC systems, which will be briefly introduced below.

Fig. 1(a) shows the conventional five-level single-rating

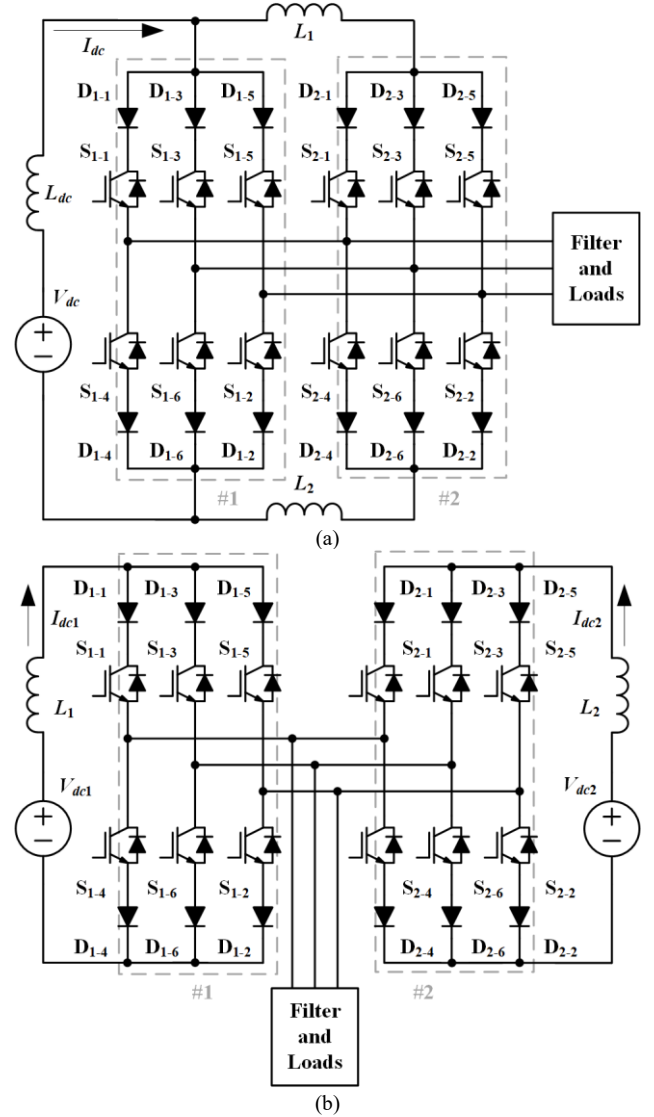


Fig. 2. Three-phase five-level CSI topologies derived from the single-rating MCSI shown in Fig. 1(a): (a) with two multi-rating inductors and (b) using two-parallel H-bridge inverters.

inductor MCSI [9], [10], which consists of two H6 converter modules connected in parallel on the AC side. In this way, five-level output currents can be obtained. On the DC side, two H6 modules are connected to the DC rail four inductors (two for each) as illustrated in Fig. 1(a). The inductors are of the same current rating, contributing to the reduction of the current ripples. However, due to the parasitic parameters in the system, the output current may be unbalanced, or the DC current may circulate between two modules. This becomes the major drawback of single-rating inductor MCSI, which hinders its applications. Notably, the single-rating inductors ( $L_{1-1}$ ,  $L_{1-2}$ ,  $L_{2-1}$ , and  $L_{2-2}$ ) can be replaced with interphase inductors (coupled inductors) as shown in Fig. 1(b). Doing so further contributes to the reduction of current ripples and overall system volume. More specifically, when the SVM is adopted to control the MCSI, the inductance of the single-rating inductors can theoretically be reduced to  $L'$  as [15]:

$$L' = (1 - \frac{1}{\sqrt{3}m_a})L_o \quad (1)$$

where  $m_a$  is the modulation index and  $L_o$  is the original inductor value.

The multi-rating inductor five-level CSI and the paralleled

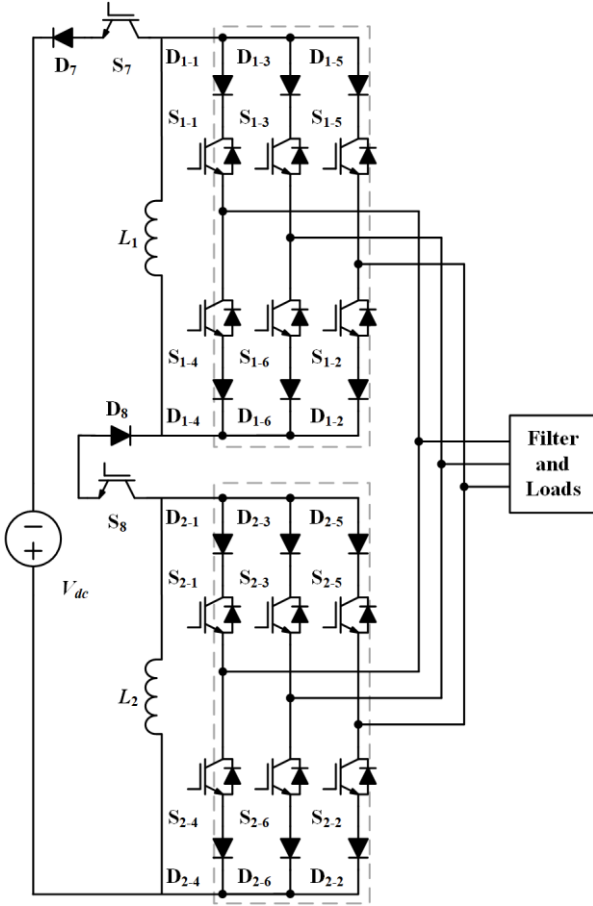


Fig. 3. Circuit schematic of the buck-boost five-level CSI.

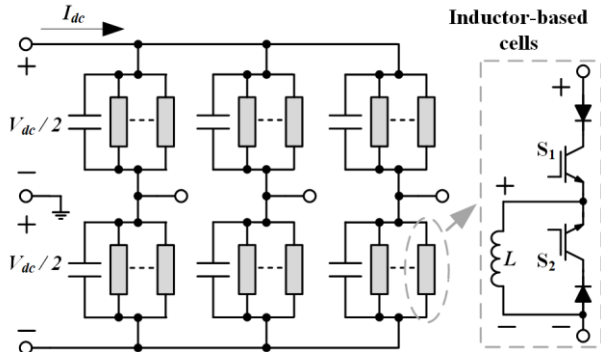


Fig. 4. Three-phase current source MMC topology with its inductor-based cells.

H-bridge five-level CSI are shown in Fig. 2(a) and (b), respectively. Both topologies can be derived from the classic single-rating inductor MCSI shown in Fig. 1(a). Clearly, each of the two five-level CSI topologies consists of two H6 CSI modules. The multi-rating inductor five-level CSI can be taken as a dual FC voltage-fed multilevel converter. Two different inductors are employed to split the input current [18], and there are no passive components in the sub-H6 modules, as shown in Fig. 2(a). However, the problem of circulating and unbalance currents still exists. To solve this, the paralleled H-bridge MCSI with two independent current sources can be adopted [19]. It resembles the single-rating inductor MCSI by using two DC current sources in such a way to address the unbalance and current-circulating issues. However, the parallel H-bridge inverter shown in Fig. 2(b) has two major disadvantages, i.e., requiring two sources, which limits practical applications, and unequal input currents, which challenges the implementation.

Furthermore, the buck-boost five-level CSI was presented to

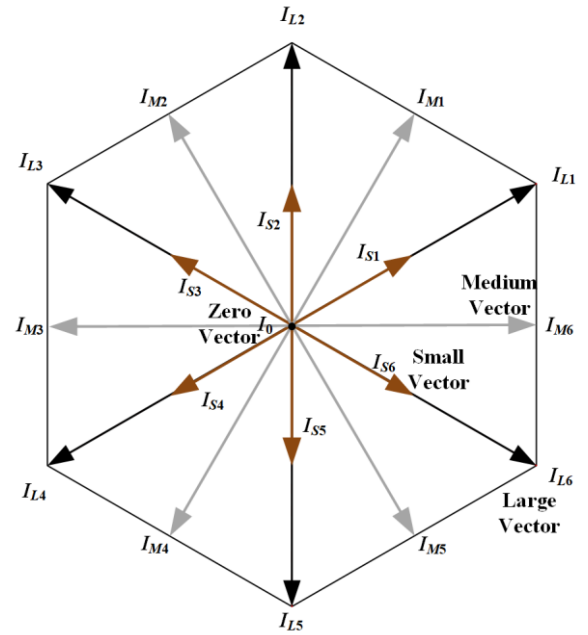


Fig. 5. Space vector diagram of the conventional five-level CSI topologies.

extend the operating range [16]. The configuration of this topology is shown in Fig. 3. As it can be observed, different from the conventional boost operation, the aim of this topology is to realize the low voltage output. However, the current unbalance issue remains, and the configuration of power switches in series with the voltage source is not recommended for high power applications, limiting the development of this MCSI topology. Another MCSI topology is the CS-MMC topology [17], which can be considered as the dual topology of the voltage-source MMC. Here, the CS-MMC employs inductor-based cells, as shown in Fig. 4, which are connected in parallel. Apparently, the CS-MMC possesses the high-power capability with integrated multiple cells, but it utilizes much more power devices and inductors, which may lead to high costs as well as high volume.

Except for the CS-MMC solution, all the above prior-art three-phase five-level CSIs consist of two H6 CSI modules. In addition, the current-available technology of the five-level SVM schemes for these MCSI topologies is mainly to control the switching combinations of two H6 converters. Normally, the space vectors can be classified into four types, i.e., zero, small, medium, and large vectors [14], as shown in Fig. 5. Then, the modulation scheme has 81 combinations for five-level CSI systems [15], which makes it very complicated and difficult to control and implement. A list of the switching combinations for the conventional five-level CSI SVM is given in Table I. The switching combinations are represented here with {xx;yy}, where the numbers of 'xx' and 'yy' refer to the corresponding ON-switch in the first and the second H6 inverter module, respectively. For instance, the vector {16; 23} means that the power devices of #1 and #6 of the first H6 inverter and the power devices of #2 and #3 of the second H6 inverter are switched on.

In fact, the conventional H6 CSI utilizes the least number of power devices, but the resultant current stress is the highest. Furthermore, the efficiency and power quality issue have been plagued for many years in the applications of CSI topologies. On the other hand, the majority of three-phase MCSI topologies use multiple H6 CSI modules to solve the above problem. However, in this case, the hardware costs as well as

TABLE I  
SWITCHING COMBINATIONS FOR THE CONVENTIONAL FIVE-LEVEL CSI

Space vectors		ON-switching combinations	Output currents		
			Phase-A	Phase-B	Phase-C
Large vectors	$I_{L1}$	{12;12}	$I_{dc}$	0	$-I_{dc}$
	$I_{L2}$	{23;23}	0	$I_{dc}$	$-I_{dc}$
	$I_{L3}$	{34;34}	$-I_{dc}$	$I_{dc}$	0
	$I_{L4}$	{45;45}	$-I_{dc}$	0	$I_{dc}$
	$I_{L5}$	{56;56}	0	$-I_{dc}$	$I_{dc}$
	$I_{L6}$	{16;16}	$I_{dc}$	$-I_{dc}$	0
Medium vectors	$I_{M1}$	{23;12} {12;23}	$0.5I_{dc}$	$0.5I_{dc}$	$-I_{dc}$
	$I_{M2}$	{34;23} {23;34}	$-0.5I_{dc}$	$I_{dc}$	$-0.5I_{dc}$
	$I_{M3}$	{45;34} {34;45}	$-I_{dc}$	$0.5I_{dc}$	$0.5I_{dc}$
	$I_{M4}$	{56;45} {45;56}	$-0.5I_{dc}$	$-0.5I_{dc}$	$I_{dc}$
	$I_{M5}$	{16;56} {56;16}	$0.5I_{dc}$	$-I_{dc}$	$0.5I_{dc}$
	$I_{M6}$	{12;16} {16;12}	$I_{dc}$	$-0.5I_{dc}$	$-0.5I_{dc}$
Small vectors	$I_{S1}$	{16;23} {23;16} {12;14} {14;12} {12;36} {36;12} {12;25} {25;12}	$0.5I_{dc}$	0	$-0.5I_{dc}$
	$I_{S2}$	{12;34} {34;12} {23;14} {14;23} {23;36} {36;23} {23;25} {25;23}	0	$0.5I_{dc}$	$-0.5I_{dc}$
	$I_{S3}$	{23;45} {45;23} {34;14} {14;34} {34;36} {36;34} {34;25} {25;34}	$-0.5I_{dc}$	$0.5I_{dc}$	0
	$I_{S4}$	{34;56} {56;34} {45;14} {14;45} {45;36} {36;45} {45;25} {25;45}	$-0.5I_{dc}$	0	$0.5I_{dc}$
	$I_{S5}$	{45;16} {16;45} {56;14} {14;56} {56;36} {36;56} {56;25} {25;56}	0	$-0.5I_{dc}$	$0.5I_{dc}$
	$I_{S6}$	{56;12} {12;56} {16;14} {14;16} {16;36} {36;16} {16;25} {25;16}	$0.5I_{dc}$	$-0.5I_{dc}$	0
Zero vector	$I_0$	{14;14} {14;36} {14;25} {36;14} {36;36} {36;25} {25;14} {25;36} {25;25} {12;45} {45;12} {23;56} {56;23} {34;16} {16;34}	0	0	0

the space occupation will be increased significantly, leading to lower power density, and also the associated issues like the current unbalance and the current-circulating are even challenging and difficult to tackle in those topologies. It is thus necessary to develop cost-effective CSI solutions.

### III. THE PROPOSED EIGHT-SWITCH FIVE LEVEL CSI

#### A. Eight-switch five-level CSI topology

In light of the above considerations, an eight-switch three-phase five-level CSI is proposed in this paper, whose circuit diagram is shown in Fig. 6. It can be observed that two shunt branches are connected in parallel between the current source and a conventional H6 converter. In specific, two power switches,  $S_7$  and  $S_8$ , are self-commutating devices. Diodes  $D_7$  and  $D_8$  are connected in series with  $S_7$  and  $S_8$  to ensure reliable reverse voltage-blocking. These two shunt-connected switches are used to bypass the input current of the rear-end H6 circuitry. Besides, diodes  $D_9$  and  $D_{10}$  are employed to prevent the circulating current. The DC-link inductors  $L_1$  and  $L_2$  are

employed for suppressing the DC current ripples. Furthermore, in order to balance the shunt-branch currents,  $L_1$  and  $L_2$  are configured with the same rating, and the interphase inductors (coupled inductors) can also be adopted to further reduce the volume. Fig. 6(a) illustrates the proposed eight-switch five-level CSI topology with normal inductor configurations. And as seen in Fig. 6(b), an inductor  $L_d$  with the interphase inductors  $L_{t1}$  and  $L_{t2}$  is included. According to (1), if using the configuration of Fig. 6(b) to generate the same current ripple, then the total DC-side inductance value ( $L_d + L_{t1}$ ) can be reduced to  $1 - \frac{1}{\sqrt{3}m_a}$  times of the original  $L_1$  [15], [20]. Regardless of the dc inductors' configuration, this paper assumes the topology in Fig. 6(a) to elaborate the operational principle and performance of the proposed CSI.

In the proposed CSI, the input current of the rear-end inverter has three states, i.e., 0,  $0.5I_{dc}$ , and  $I_{dc}$ . Depending on the output current modes, the space vectors can be classified as zero, small, and large vectors. Accordingly, there will be 37 switching combinations in theory for the whole converter,

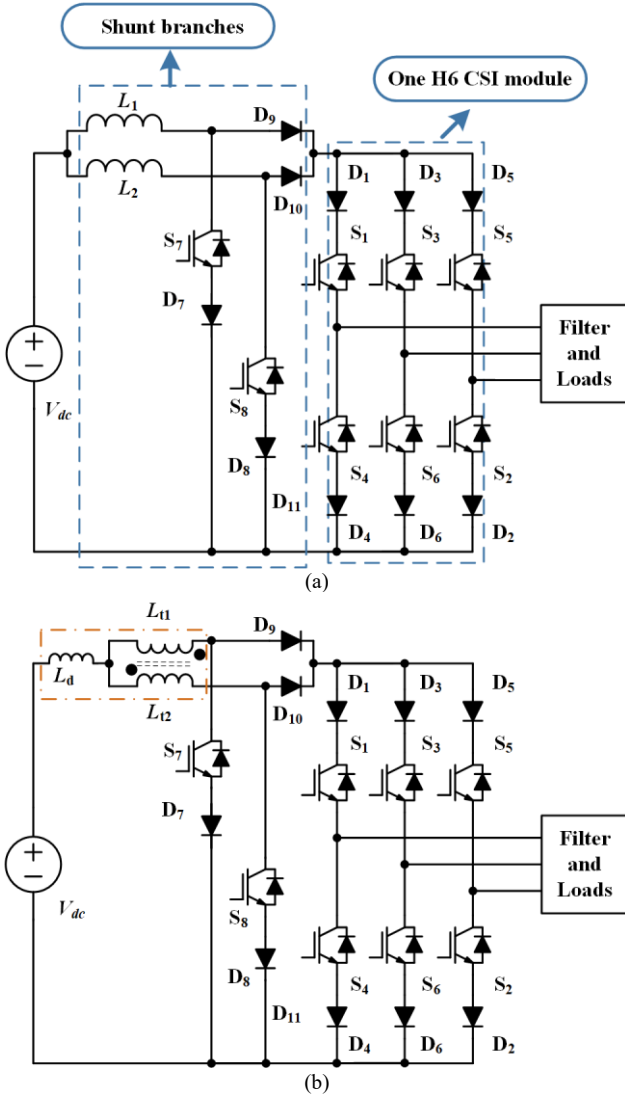


Fig. 6. Proposed eight-switch five-level CSI topology with (a) two same rating inductors and (b) interleaved inductors.

including 6 large-vector switching combinations, 12 small-vector switching combinations and 19 zero-vector switching combinations. All the possible switching combinations with the corresponding output currents are categorized in Table II, where the numbers refer to the corresponding ON-switch in the proposed CSI. However, it can be found that most of the zero-vector switching combinations are redundant, for example, {1478} gives two paths for shunt-DC currents. In order to obtain lower conduction currents as well as the dissipation, the zero vector can select with the switching combination of {78}. Therefore, when excluding the redundant switching combinations and considering the discharging period of the two inductors, the modulation states of the proposed eight-switch five-level CSI can be defined into 13 vectors, which can then be obtained with 19 possible switching combinations.

Fig. 7 further shows the exemplified equivalent circuits for each kind of vectors. In detail, the zero vector can be obtained, when both  $S_7$  and  $S_8$  are turned ON, and then all the output currents are bypassed as shown in Fig. 7(a). When only  $S_7$  or  $S_8$  is turned ON, a small vector can be generated as shown in Fig. (b), and in this case, the output current will be  $0.5I_{dc}$ . When  $S_7$  and  $S_8$  are both turned OFF as shown in Fig. 7(c), the large vectors can be generated being the same as those for the conventional H6 CSI, where only one upper-arm switch ( $S_1, S_3,$

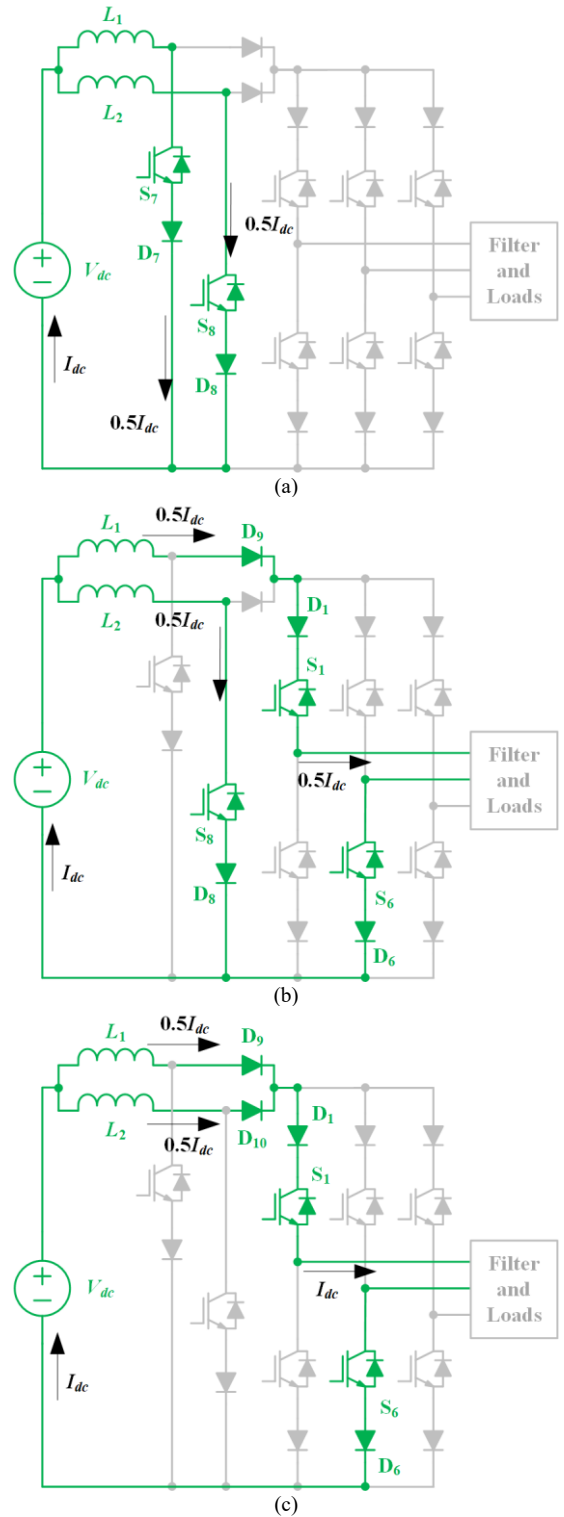


Fig. 7. Exemplified equivalent circuits of different modulation states: (a) zero vector, (b) small vector, and (c) large vector.

$S_5$ ) and one-lower arm switch ( $S_4, S_6, S_2$ ) of the rear-end H6 inverter are turned ON, leading to the output current being  $I_{dc}$ . Therefore, the proposed CSI can generate a five-level output current.

Compared to the prior-art three-phase five-level CSIs, one feature of the proposed CSI topology is that it utilizes fewer power switching devices. That is, the eight-switch CSI topology does not require two H6 converters (i.e., in total, a minimum of 12 power switches and 12 power diodes), but eight power switches and 10 power diodes to generate a five-level output. In addition, the SVM method for the

TABLE II  
SWITCHING COMBINATIONS FOR THE PROPOSED FIVE-LEVEL CSI

Space vectors		ON-switching combinations	Output currents		
			Phase-A	Phase-B	Phase-C
Large vectors	$I_{L1}$	{12}	$I_{dc}$	0	$-I_{dc}$
	$I_{L2}$	{23}	0	$I_{dc}$	$-I_{dc}$
	$I_{L3}$	{34}	$-I_{dc}$	$I_{dc}$	0
	$I_{L4}$	{45}	$-I_{dc}$	0	$I_{dc}$
	$I_{L5}$	{56}	0	$-I_{dc}$	$I_{dc}$
	$I_{L6}$	{16}	$I_{dc}$	$-I_{dc}$	0
Small vectors	$I_{S1}$	{127} {128}	$0.5I_{dc}$	0	$-0.5I_{dc}$
	$I_{S2}$	{237} {238}	0	$0.5I_{dc}$	$-0.5I_{dc}$
	$I_{S3}$	{347} {348}	$-0.5I_{dc}$	$0.5I_{dc}$	0
	$I_{S4}$	{457} {458}	$-0.5I_{dc}$	0	$0.5I_{dc}$
	$I_{S5}$	{567} {568}	0	$-0.5I_{dc}$	$0.5I_{dc}$
	$I_{S6}$	{167} {168}	$0.5I_{dc}$	$-0.5I_{dc}$	0
Zero vector	$I_0$	{14} {36} {25} {78} {147} {367} {257} {148} {368} {258} {1478} {3678} {2578} {1278} {2378} {3478} {4578} {5678} {1678}	0	0	0

proposed CSI can be much simpler, which avoids the complicated combinations of the dual converters.

### B. Space Vector Implementation

As mentioned above and listed in Table II, there are 13 potentially available space vectors that can be obtained by 19 switching combinations, and subsequently, five-level output currents can be generated. In order to generate a proper switching sequence, the modulation should follow these principles:

- Always use the nearest current vectors to generate the reference vector to reduce output harmonics.
- Keep the switching operation of  $S_7$  and  $S_8$  equally distributed in one switching period in order to reduce the dc inductors' current ripple.
- Use  $S_7$  and  $S_8$  to add certain special modulation-state segments to ensure the switching instants completed under lower current stresses and lower power dissipation.

To select and use the nearest vectors to synthesize the reference vector, the general method is to use three vectors to define the triangular region, in which the target reference vector locates. Fig. 8 illustrates this synthesizing principle. If let any three vectors ( $\vec{i}_a$ ,  $\vec{i}_b$  and  $\vec{i}_c$ ) to define a triangular region in the  $\alpha\beta$  plane, and the reference vector  $\vec{i}_{ref}$  locates within this region, then  $\vec{i}_{ref}$  can be synthesized from

$$\vec{i}_{ref} = \vec{k}_a + \vec{k}_b + \vec{i}_c = \frac{|\vec{k}_a|}{n_a} \vec{i}_a + \frac{|\vec{k}_b|}{n_b} \vec{i}_b + (1 - \frac{|\vec{k}_a|}{n_a} - \frac{|\vec{k}_b|}{n_b}) \vec{i}_c \quad (2)$$

where  $n_a$  and  $n_b$  are the distances between the adjacent vectors.

As exemplified in Fig. 9(a) for Sector I, each sextant has 2

large vectors, 2 small vectors and 1 zero vector. Based on the above principles, the modulation scheme is proposed to partition each sector into five regions. That is, in this paper, the overall space vector diagram of the eight-switch five-level CSI can be divided into 30 regions, as shown in Fig. 9(b).

The modulation strategy can be classified into two operational modes: Mode 1 (three-level switched currents) and Mode 2 (five-level switched currents), which will be illustrated below in detail by taking Fig. 9(a) as an example.

In Mode 1, the reference vector locates in Region 1 ( $\Delta ABE$ ), which is defined within  $\{I_{S6}, I_{S1}, I_0\}$ , as shown in Fig. 9(a). That is, the converter modulates with the same zero and small vectors as the conventional five-level CSI solutions. And the AC output will be 3-level switching currents. The detailed operational sequences for Sector I in one modulation period are illustrated in Fig. 10, where  $t_{act1}$  and  $t_{act2}$  are the switching transitions for the rear-end H6-inverter switches ( $S_1$ - $S_6$ ), while  $t_{n1}$ - $t_{n6}$  are the switching transitions for the shunt branch switches of  $S_7$  and  $S_8$ . In addition, the switching interval between  $t_{n2}$  and  $t_{n3}$  and the switching interval between  $t_{n4}$  and  $t_{n5}$  are generated to ensure the zero current switching (ZCS) for  $S_2$  and  $S_6$ , respectively, where  $S_7$  and  $S_8$  are turned ON simultaneously, leading to the expected zero vector. This ZCS configuration makes the equivalent commutation behave the same as that for the H7 CSI in [21]. It has been discussed and demonstrated in [21] that, the shunt connected switch can take over all the switching losses, and effectively promote the efficiency of CSI.

In mode 2, when the modulation index  $m_a$  is above 0.5, the reference in Sector I can rotate among Region 2, Region 3, Region 4, and Region 5, with 5-level switched currents at the AC output. It is defined that Region 2 and Region 5, Region 3 and Region 4 are symmetric about the angle bisector of Sector I. When the reference vector locates in Region 2, the nearest three composition space vectors are  $\{I_{L6}, I_{S1}, I_{S6}\}$ . Similarly, the vector compositions for Region 5 are  $\{I_{S6}, I_{L1}, I_{S1}\}$ , as shown in



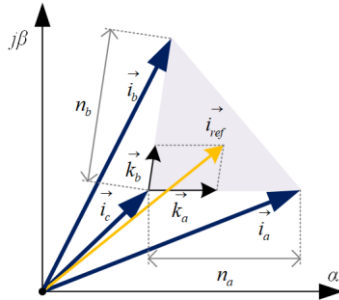
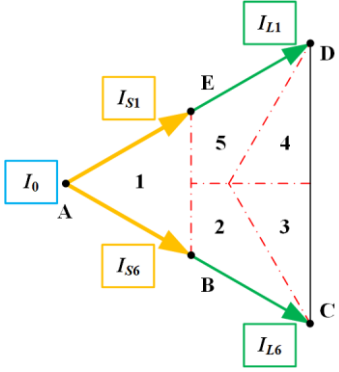
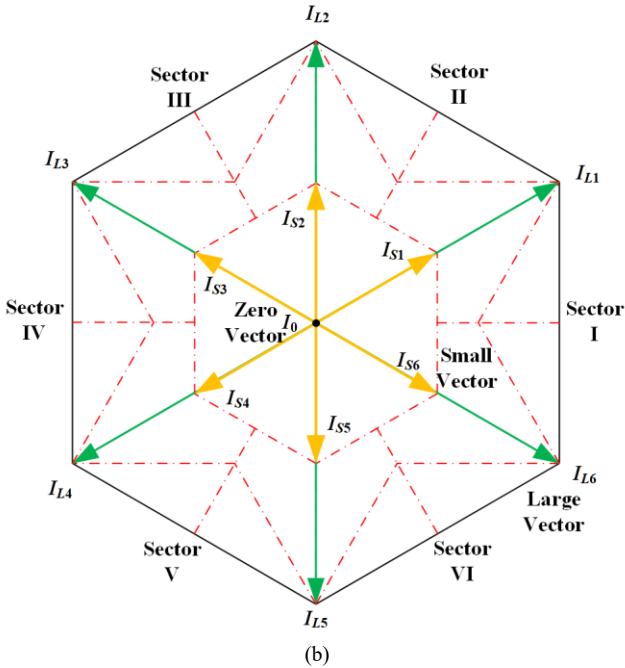


Fig. 8. Illustration of synthesizing the reference current with three arbitrary vectors.



(a)



(b)

Fig. 9. Vector diagrams of the proposed SVM for the eight-switch five-level CSI topology: (a) region divisions in Sector I and (b) overall space vectors.

Fig. 9(a). The corresponding switching intervals for Region 2 and Region 5 are demonstrated in Fig. 11(a) and Fig. 11(d), respectively. In those switching intervals,  $S_7$  or  $S_8$  is employed to keep an extended turning-ON interval within  $T_{ins}$ , so that the switching for  $S_2$  and  $S_6$  can operate with only  $0.5I_{dc}$ . As for Region 3 in Fig. 9(a), the nearest space vector composition will be  $\{I_{L6}, I_{L1}, I_{S6}\}$ . However, in order to switch  $S_2$  and  $S_6$  with small vector ( $0.5I_{dc}$ ), another small space vector  $I_{S1}$  should be inserted. That is, the vector compositions become  $\{I_{S6}, I_{L6}, I_{S1}, I_{L1}\}$ , and the switching sequences are demonstrated in Fig. 11(b). Accordingly, the vector compositions of Region 4 are the same, i.e.,  $\{I_{L6}, I_{S1}, I_{L1}, I_{S6}\}$  but with different dwell time and switching sequence, and the switching sequences are demonstrated in Fig. 11(c). It is noted that region 3 and region

Region 1 in Sector I

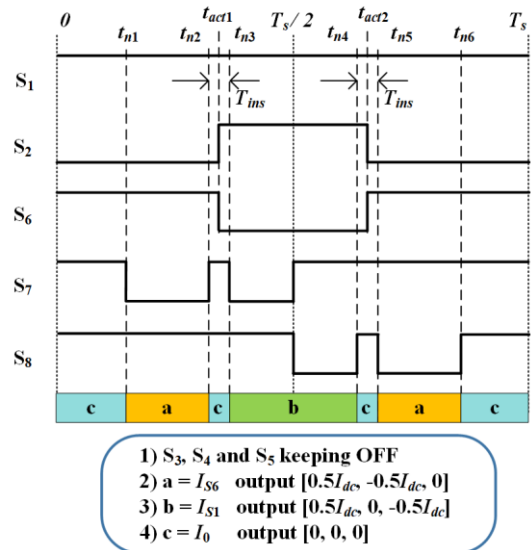


Fig. 10. Switching interval operation of Mode 1 (Region 1, in one period).

4 should have the additional intervals ( $T_{ins}$ ) for  $S_7$  or  $S_8$  to ensure the switching of  $S_2$  and  $S_6$  being under a lower current stress, i.e.,  $0.5I_{dc}$ .

It should be noted that there is no medium vector (e.g.,  $\{12;16\}$  of the conventional five-level SVM in Table I) in the proposed five-level CSI modulation because the current can only flow through two phase legs in one converter-bridge. In addition, for balancing the inductor currents and keeping low current ripples, the power switches  $S_7$  and  $S_8$  should be operated with an equal duration in one switching period. Therefore, in the proposed modulation scheme, the switching sequences of  $S_7$  are theoretically in symmetry with that of  $S_8$  in a half-period.

According to the above modulation scheme, the dwell time of the SVM states should be carefully calculated. Taking Region 2 of Sector I as an example, the current reference should be synthesized with  $I_{L6}$ ,  $I_{S1}$  and  $I_{S6}$  referring to Fig. 9(a), which are expressed as

$$\begin{cases} I_{L6} = \frac{2\sqrt{3}}{3} I_{dc} e^{-j\frac{\pi}{6}} \\ I_{S1} = \frac{\sqrt{3}}{3} I_{dc} e^{j\frac{\pi}{6}} \\ I_{S6} = \frac{\sqrt{3}}{3} I_{dc} e^{-j\frac{\pi}{6}} \end{cases} \quad (3)$$

Subsequently, the reference vector  $I_{ref}$  can be synthesized using

$$\begin{cases} I_{ref} = m_a I_{dc} e^{j\theta} \\ T_s I_{ref} = T_a I_{L6} + T_b I_{S1} + T_c I_{S6} \\ T_a + T_b + T_c = T_s \end{cases} \quad (4)$$

where  $m_a$  is the modulation index,  $T_s$  is the switching period, and  $T_a$ ,  $T_b$ , and  $T_c$  refer to the corresponding dwell time within one period for vectors  $I_{L6}$ ,  $I_{S1}$  and  $I_{S6}$ .

As for Region 3 of Sector I, the current reference is synthesized with four vectors  $\{I_{S6}, I_{L6}, I_{S1}, I_{L1}\}$  as shown in Fig. 9(a), among of which, current vectors  $I_{S6}$ ,  $I_{L6}$ , and  $I_{S1}$  are the same as (3), and the vector  $I_{L1}$  can be expressed as



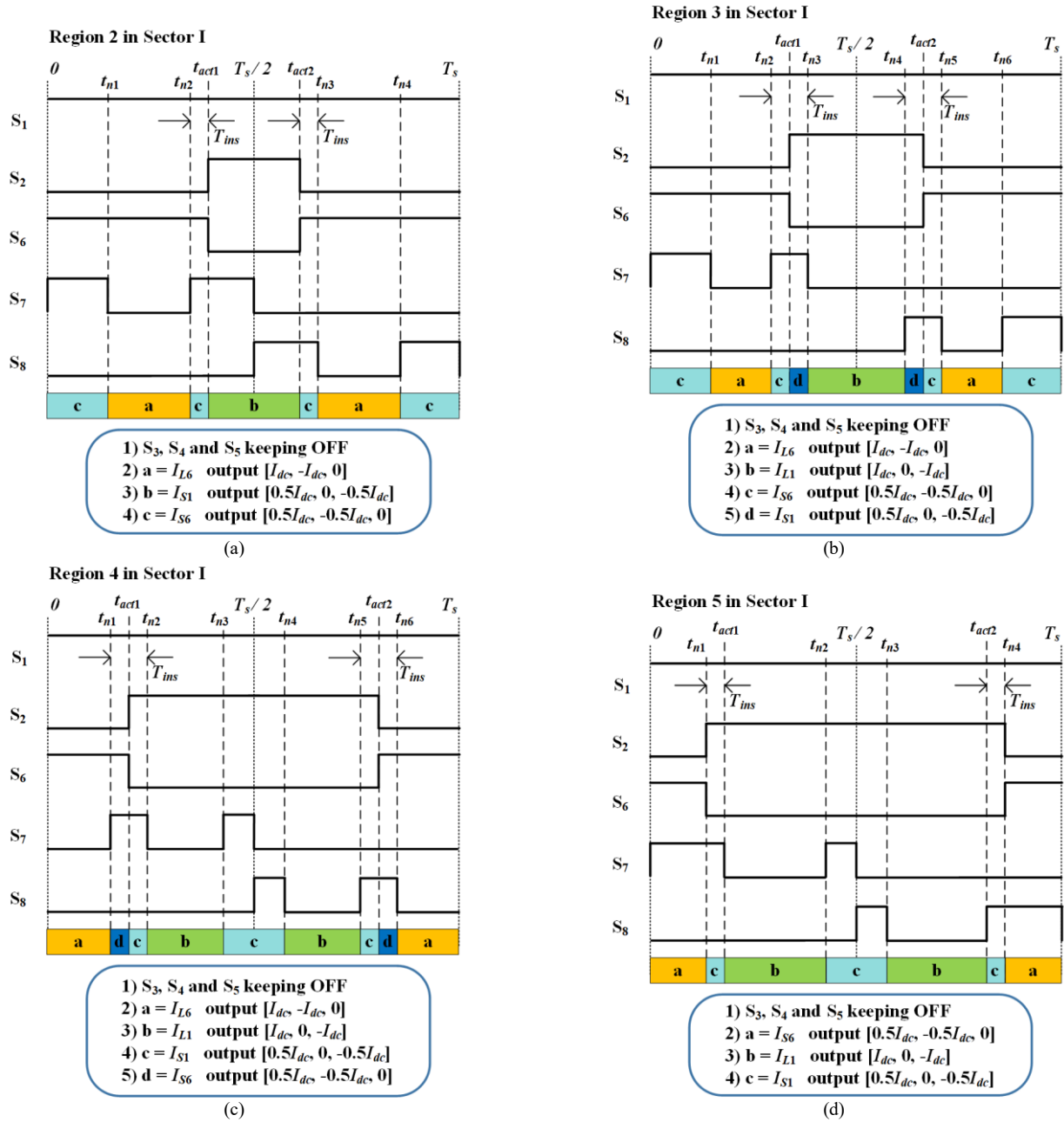


Fig. 11. Switching interval operation of Mode 2 in one period: (a) Region 2, (b) Region 3, (c) Region 4, and (d) Region 5.

$$I_{L1} = \frac{2\sqrt{3}}{3} I_{dc} e^{j\frac{\pi}{6}} \quad (5)$$

The state  $I_{S1}$  is the pre-set interval to guarantee the switching completion of  $S_2$  and  $S_6$ . If the dwell time of  $I_{S1}$  is denoted as  $T_d$ , the current reference vector can be synthesized with

$$\begin{cases} I_{ref} = m_a I_{dc} e^{j\theta} \\ T_s I_{ref} = T_a I_{L6} + T_b I_{L1} + T_c I_{S6} + T_d I_{S1} \\ T_a + T_b + T_c + T_d = T_s \end{cases} \quad (6)$$

Specifically,  $T_d$  can be regulated by  $T_{ins}$ , following

$$T_{[n2,act1]} = T_{[act1,n3]} = T_{[n4,act2]} = T_{[act2,n5]} = \frac{1}{2} T_{ins} = \frac{1}{2} T_d \quad (7)$$

where  $T_a$ ,  $T_b$ ,  $T_c$ , and  $T_d$  refer to the corresponding dwell time within one period for vectors  $I_{L6}$ ,  $I_{L1}$ ,  $I_{S6}$  and  $I_{S1}$ , and  $T_{[n2,act1]}$ ,  $T_{[act1,n3]}$ ,  $T_{[n4,act2]}$ , and  $T_{[act2,n5]}$  are the corresponding time

interval between each switching transitions. In a similar way, all dwell time can be calculated, and the resultant dwell time in Sector I has been summarized in Table III.

### C. Current Balancing Scheme

The current unbalance is a typical problem for multi-level CSIs. In fact, the unbalanced current input will degrade the output power quality, and may result in instability or even system damage [22]-[26]. Thus, a few current balancing methods for the single-rating inductor MCSI have been presented in the literature. Some of these current balancing solutions modified the modulation strategies, e.g., the phase-shifted PWM [24], but with a basic open-loop control, which cannot guarantee the accuracy. For a closed-loop scheme, the common method for current balancing is to sample the DC input currents (such as  $I_{dc1-1}$  and  $I_{dc2-1}$  in Fig. 1(a)), and uses the sampled currents to re-distribute the vector states (zero vectors [20], active vectors [25], and medium vectors [26]) in the modulation schemes. In this way, the unbalancing

TABLE III  
DWELL TIME AND VECTORS IN SECTOR I

Region	Vector	Overall Dwell Time
1	$I_{S6}$	$T_a = 2m_a T_s \sin(\frac{\pi}{6} - \theta)$
	$I_{S1}$	$T_b = 2m_a T_s \sin(\frac{\pi}{6} + \theta)$
	$I_0$	$T_c = T_s - T_a - T_b$
2	$I_{L6}$	$T_a = T_s (2m_a \cos \theta - 1)$
	$I_{S1}$	$T_b = 2m_a T_s \sin(\frac{\pi}{6} + \theta)$
	$I_{S6}$	$T_c = T_s - T_a - T_b$
3	$I_{L6}$	$T_a = T_s [\sqrt{3}m_a \sin(\frac{\pi}{3} - \theta) - 1] + \frac{1}{2}T_{ins}$
	$I_{L1}$	$T_b = m_a T_s \sin(\frac{\pi}{6} + \theta) - \frac{1}{2}T_{ins}$
	$I_{S6}$	$T_c = T_s - T_a - T_b - T_d$
	$I_{S1}$	$T_d = T_{ins}$
4	$I_{L6}$	$T_a = m_a T_s \sin(\frac{\pi}{6} - \theta) - \frac{1}{2}T_{ins}$
	$I_{L1}$	$T_b = T_s [\sqrt{3}m_a \sin(\frac{\pi}{3} + \theta) - 1] + \frac{1}{2}T_{ins}$
	$I_{S1}$	$T_c = T_s - T_a - T_b - T_d$
	$I_{S6}$	$T_d = T_{ins}$
5	$I_{S6}$	$T_a = 2m_a T_s \sin(\frac{\pi}{6} - \theta)$
	$I_{L1}$	$T_b = T_s (2m_a \cos \theta - 1)$
	$I_{S1}$	$T_c = T_s - T_a - T_b$

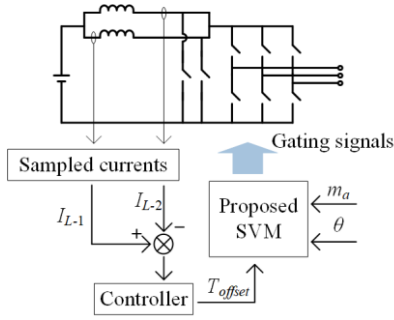


Fig. 12. Control diagram for current balancing scheme.

issue can be addressed to a large extent. However, these solutions incur heavy computation, since for the basic modulation there are already 81 switching combinations.

Obviously, the proposed modulation scheme ensures the power switches  $S_7$  and  $S_8$  with symmetrical equal operation to balance the inductor currents  $I_{L-1}$  and  $I_{L-2}$ . In addition, the inductors  $L_1$  and  $L_2$  are configured with the same current rating. However, in practical applications, the currents of  $I_{L-1}$  and  $I_{L-2}$  may not be exactly the same, because of the electrical parasitic parameters of the two shunt branch components. Therefore, the current balancing issue should also be considered in the proposed topology.

Nevertheless, the DC side configuration of the proposed CSI topology has some unique features. That is,  $I_{L-1}$  and  $I_{L-2}$  are mainly controlled by  $S_7$  and  $S_8$  and their turn-on operation will charge  $L_1$  and  $L_2$ , respectively. Fortunately,  $S_7$  and  $S_8$  theoretically have an equivalent role for the rear-end inverter, so that the switching of  $S_7$  or  $S_8$  can be controlled independently as long as the sum of their dwell time in one modulation period is a certain fixed value. Therefore, considering the above features, the basic operation for the proposed CSI topology is to adjust the switching transitions of

$S_7$  and  $S_8$  in a way to balance the inductor currents, while keeping the same overall dwell time of each state.

When the switch  $S_7$  (or  $S_8$ ) is turned ON, the inductor  $L_1$  (or  $L_2$ ) will be directly charged by the DC source  $V_{dc}$ . Therefore, the peak and trough values of the two inductor current ripples are interleaved by a half switching period. Setting the inductor currents to be sampled every half switching cycle can obtain the initial current  $I_{z1}$ ,  $I_{z2}$  and the half cycle current  $I_{h1}$ ,  $I_{h2}$ . And let  $I_{L-1}$  and  $I_{L-2}$  be the actual sampled average currents of the two inductors and  $I_{L-1}^*$  and  $I_{L-2}^*$  be the demanded ideal current values, the corresponding currents can be expressed as (8) when assuming  $I_{L-1}$  is larger than  $I_{L-2}$ .

$$\begin{cases} I_{L-1} = \frac{1}{2}(I_{z1} + I_{h1}) \\ I_{L-2} = \frac{1}{2}(I_{z2} + I_{h2}) \\ \Delta I_{L-1} = I_{L-1} - I_{L-1}^* = \frac{V_{dc}}{L_1} \frac{1}{2} T_{offset} \\ \Delta I_{L-2} = I_{L-2}^* - I_{L-2} = \frac{V_{dc}}{L_2} \frac{1}{2} T_{offset} \\ I_{L-1}^* = I_{L-2}^* = \frac{1}{2} I_{dc} \end{cases} \quad (8)$$

where  $T_{offset}$  is the switching transition offset. Depending on the above equations,  $T_{offset}$  can be expressed as

$$T_{offset} = \frac{2(I_{L-1} - I_{L-2})}{V_{dc}} \frac{L_1 L_2}{L_1 + L_2} \quad (9)$$

Therefore, the operation of  $S_7$  and  $S_8$  can be adjusted by  $T_{offset}$  in order to obtain the relatively equal inductor currents. Taking Region 2 of Sector I as an example (shown in Fig. 11(a)), the switching transitions  $t_{n1}$  and  $t_{n4}$  can be modified as

$$\begin{cases} t_{n1}' = t_{n1} - T_{offset} \\ t_{n4}' = T_s - t_{n1}' - 2T_{offset} \end{cases} \quad (10)$$

where  $t_{n1}$  and  $t_{n4}$  are the original switching transitions, and  $t_{n1}'$  and  $t_{n4}'$  refer to the regulated new transitions. Notably, other operational intervals remain unchanged in this case.

In a similar way, this switching regulation scheme can be applied to all the modulation regions, and the resultant inductor currents will then be balanced. Fig. 12 shows the control diagram of the proposed current balancing scheme. The inductor currents are sampled by half switching cycle to obtain the average values  $I_{L-1}$  and  $I_{L-2}$ . And then, they are subtracted and the result is given into the controller to produce the switching transition offset  $T_{offset}$  using (8)-(9). With the information of modulation index  $m_a$  and delay angle  $\theta$ , the proposed modulation strategy is able to calculate out the actual modified switching transitions, which can effectively resolve the current balancing issue.

#### D. Operational feature benchmarking

Table IV compares the switching and conducting characteristics in Sector I for one carrier period (with the symmetric 5-segment modulation) among the four selected MCSI solutions, i.e., the conventional H6 CSI, conventional 5-level CSI, and the proposed eight-switch 5-level CSI.

The power-device count of the proposed eight-switch five-level CSI is obviously less than the conventional

TABLE IV  
COMPARISON OF SWITCHING AND CONDUCTING CHARACTERISTICS  
IN SECTOR I FOR ONE CARRIER PERIOD (WITH SYMMETRIC 5-SEGMENT MODULATION)

Topology	Conventional H6 CSI	Single-rating inductor 5-level CSI	Proposed eight-switch 5-level CSI			
Output waveform	3-level	5-level	5-level			
Power Switches	6	12	8			
Power diodes	6	12	10			
Switching current stress	$S_1-S_6$	$S_1-S_{12}$	Mode 1		Mode 2	
	$I_{dc}$	$0.5I_{dc}$	$S_1-S_6$	$S_7, S_8$	$S_1-S_6$	$S_7, S_8$
Switching counts	8	16	ZCS	$0.5I_{dc}$	$0.5I_{dc}$	$0.5I_{dc}$
Conducting current	$I_{dc}$	$0.5I_{dc}$	Zero	Small	Large	
			$0.5I_{dc}$	$0.5I_{dc}$	$I_{dc}$	
Conducting switches	2	4	2	3	2	
Conducting diodes	2	4	2	4	$2(0.5I_{dc})+2(I_{dc})$	

single-rating five-level CSI. More specifically, the switch count has reduced by one third. The proposed CSI has two operational modes. In Mode 1, switches  $S_1-S_6$  can operate with zero current switching, while  $S_7$  and  $S_8$  switch with  $0.5I_{dc}$  resulting in 12 switching counts per switching cycle. In Mode 2, all the power switches can operate with  $0.5I_{dc}$ , and the total switching counts is 12 either. It means that the switching losses of the proposed CSI can be three-quarters of the conventional single-rating five-level CSI.

For the current stress, the current rating through the additional switch  $S_7$  or  $S_8$  is half of that in the rear-end switches ( $S_1-S_6$ ) for the proposed topology as summarized in Table IV, which means the switches  $S_1-S_6$  (together with  $D_1-D_6$ ) can be relatively high power but cheap devices, while  $S_7$  and  $S_8$  (may together with  $D_7-D_{10}$ ) can be employed by high

performance devices with low current rating. This feature enables a cost-effective customization according to the CSI applications. Furthermore, it gives the possibility to improve the efficiency by only using SiC switches for  $S_7$  and  $S_8$ .

#### IV. PERFORMANCE EVALUATION

##### A. Simulation Results

Referring to Fig. 6(a), a simulation model has been built up in MATLAB/SIMULINK to evaluate the performance of the proposed eight-switch five-level CSI topology. The system parameters are listed in Table V. The DC current is maintained at 12 A in the simulations. Resistor of  $16 \Omega$  is adopted as the load with AC output capacitor filter. The performance of the proposed MCSI is compared with the conventional H6 inverter.

Fig. 13 first compares the phase-A output currents of the conventional H6 CSI and the proposed eight-switch five-level CSI, where the modulation index  $m_a$  is 0.8. In this case, the H6 CSI is employed with a 5-mH DC inductor to obtain a relatively comparable DC inductor current ripple, while other configuration parameters are the same as the proposed converter. It can be seen in Fig. 13 that the proposed eight-switch CSI topology achieves a five-level output current (blue lines) as expected, while the H6 CSI only produces three-level switching current. When the AC capacitor filters are added, the corresponding filtered currents of phase-A are shown as the red lines. Furthermore, the three-phase output voltages and currents of the proposed CSI are shown in Fig. 14. The performance of the proposed MCSI is further validated

TABLE V  
PARAMETERS OF THE PROPOSED EIGHT-SWITCH  
FIVE-LEVEL CURRENT SOURCE INVERTER

Parameters	Value
Power rating	3.18 kW
DC current $I_{dc}$	12 A
DC inductance, $L_1, L_2$	5 mH
AC filter capacitance	10 $\mu$ F
Switching frequency ( $S_1-S_6$ )	5 kHz
Switching frequency ( $S_7, S_8$ )	10 kHz
Output AC frequency	50 Hz
Load resistance	16 $\Omega$

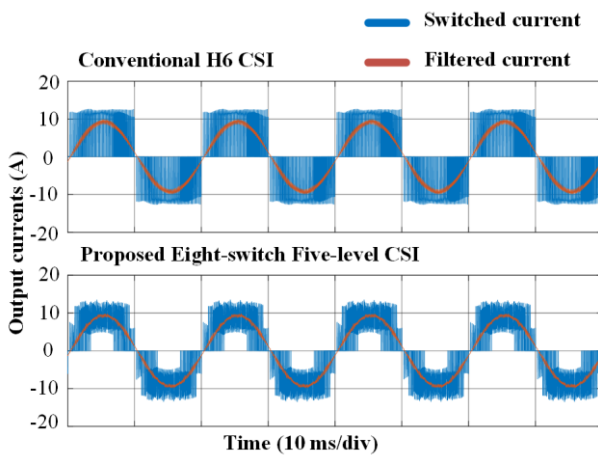


Fig. 13. Simulation results of the proposed converter (output currents of phase-A).

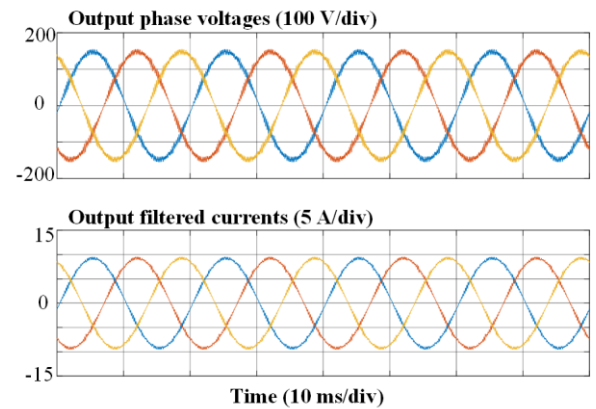


Fig. 14. Output three-phase voltages and filtered currents of the proposed five-level CSI.

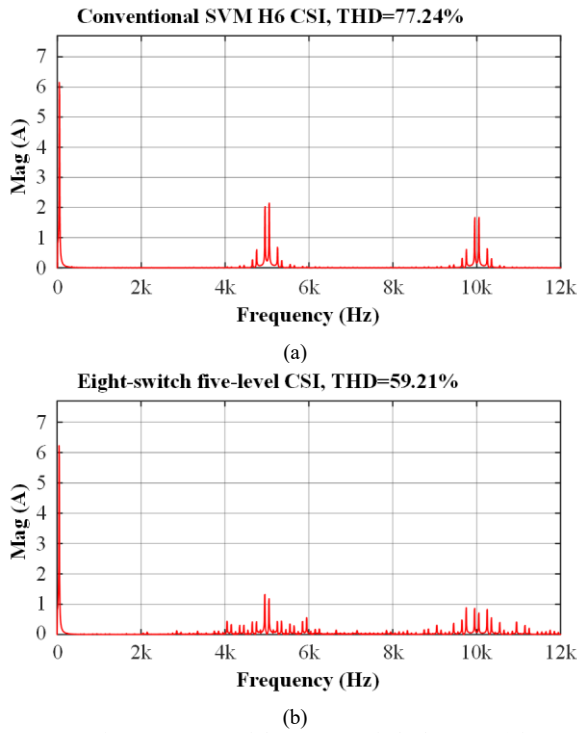


Fig. 15. Harmonic components of the output switched current (Fig. 13) of (a) the conventional H6 CSI and (b) the proposed eight-switch five-level CSI.

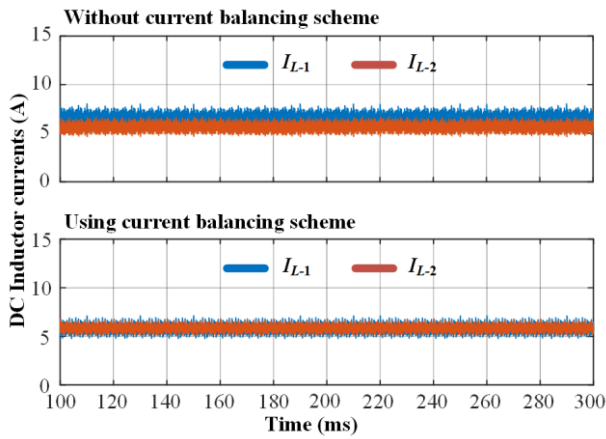


Fig. 16. Comparisons (simulations) of the inductor currents of the eight-switch five-level CSI topology without and with the current balancing control scheme ( $L_1$ : 4.5 mH,  $L_2$ : 5.5 mH).

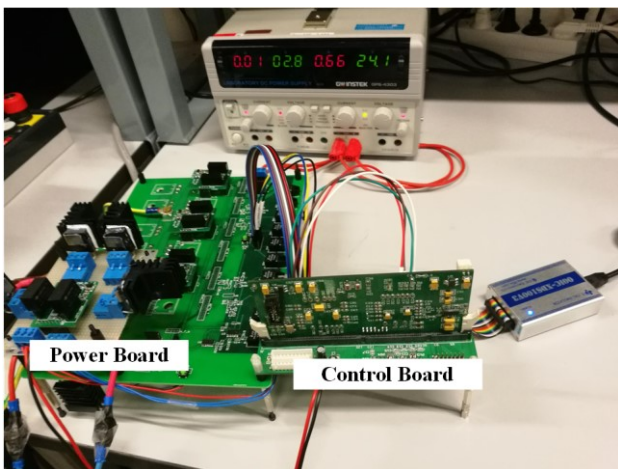


Fig. 17. Experimental setup of the proposed eight-switch five-level CSI.

through the fast Fourier transform (FFT) analysis of the output switching currents (see Fig. 13), as shown in Fig. 15, where the

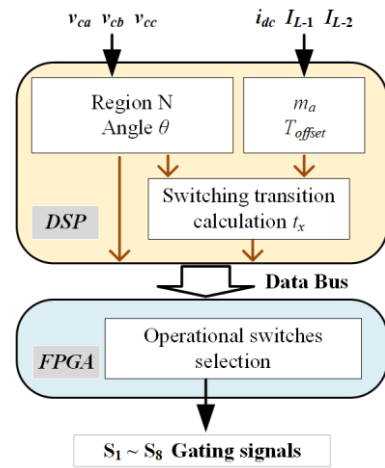


Fig. 18. Implemented control system for the proposed eight-switch five-level CSI shown in Fig. 17.

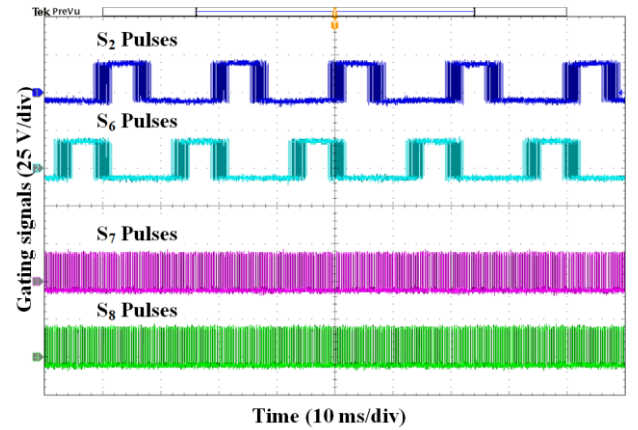


Fig. 19. Gating sequences (experiments) for the power devices  $S_2$ ,  $S_6$ ,  $S_7$  and  $S_8$  of the proposed eight-switch five-level CSI topology.

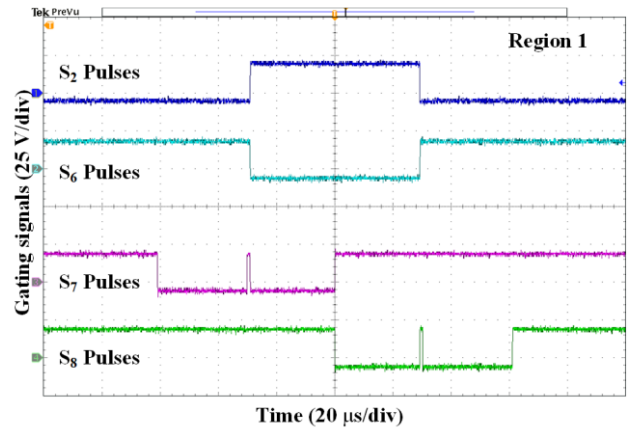


Fig. 20. Detailed switching intervals (experiments) for the power devices  $S_2$ ,  $S_6$ ,  $S_7$  and  $S_8$  of Region 1 ( $m_a$ : 0.3).

switching frequency harmonic components in the proposed CSI are much lower than the conventional H6 CSI. As a result, the total harmonic distortion (THD) value of the switching current in eight-switch five-level CSI is 59.21 %, while it is 77.24% for the conventional H6 CSI topology. Therefore, it has been validated that the proposed eight-switch five-level CSI has better output performances over the conventional H6 CSI.

Additionally, the proposed current balancing scheme is also tested. The inductance of the eight-switch five-level CSI in the simulation model is changed, where  $L_1$  is 4.5 mH and  $L_2$  is 5.5 mH. Fig. 16 shows the inductor currents of the proposed



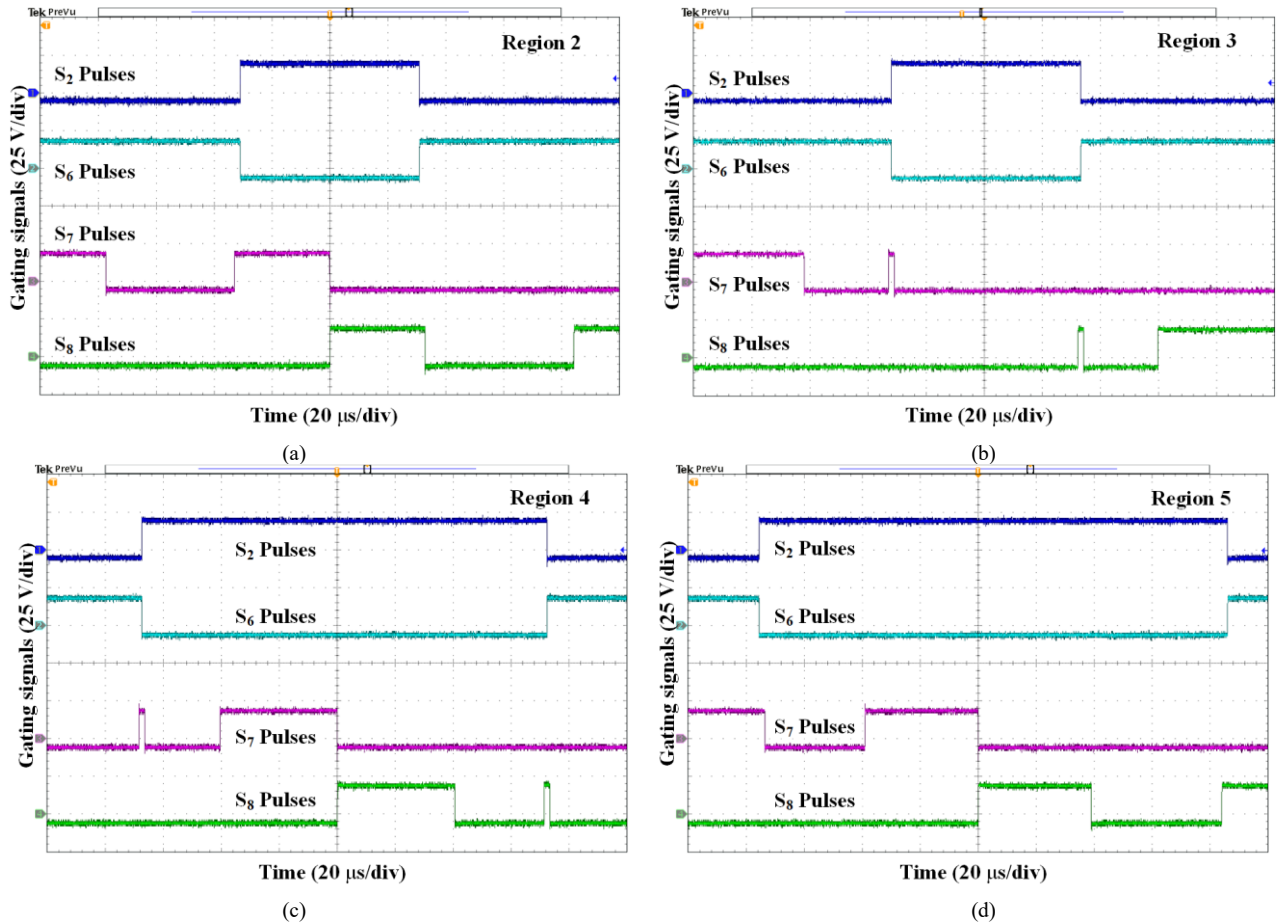


Fig. 21. Detailed switching intervals (experiments) for the power devices  $S_2$ ,  $S_6$ ,  $S_7$  and  $S_8$  of (a) Region 2, (b) Region 3, (c) Region 4, and (d) Region 5 ( $m_a$ : 0.8).

inverter without and with the current balancing control. As observed in Fig. 16, when the converter is operating without any current balancing control but only with the basic modulation scheme, the currents through  $L_1$  and  $L_2$  are not balanced, where  $L_1$  carries more input current than  $L_2$  and withstands a relatively higher current ripple. By contrast, when employing the proposed current balancing scheme discussed in Section III(C), the currents  $I_{L-1}$  and  $I_{L-2}$  of the proposed CSI topology are well balanced. The effectiveness of the current balancing strategy is thus verified.

### B. Experimental Verifications

To further validate the performance of the proposed eight-switch five-level CSI topology, a downscale experimental prototype has been built up in the laboratory. Fig. 17 shows the experimental setup, and Fig. 18 illustrates the implementation diagram of the control system, respectively. As shown in Fig. 18, on the control board, the AC voltages and the sampled DC inductor currents are input to the micro controller unit, which is a Digital Signal Processor (DSP, TMS320F28335). The region judgement and the switching time calculation (together with the current balancing algorithm) are implemented in the DSP controller. Then, the calculated results are directly transferred to another processor unit, i.e., an FPGA xc3s500e from XILINX Spartan3E, to implement the switching selection and the driving pulses generation functions. On the power board, the other power switches  $S_1$ - $S_6$  are silicon IGBT devices from Infineon (part no.: IKW20N60T), while all the power diodes are from CREE (part no.: C2D20120D). To obtain a cost-effective performance, power switches  $S_7$  and  $S_8$  (part no.: C2M0160120D) are SiC devices from CREE. The

other parameters are the same as those listed in Table V.

With the proposed SVM scheme presented in Section III, the experimental gating sequences for  $S_2$ ,  $S_6$ ,  $S_7$ , and  $S_8$  are shown in Fig. 19. Clearly, the extra power devices  $S_7$  and  $S_8$  are switched more frequently than the other power switches in the proposed modulation. However, the use of SiC power devices for  $S_7$  and  $S_8$  will not compromise the entire system efficiency. In addition, the driving pulses for switches  $S_2$ ,  $S_6$ ,  $S_7$ , and  $S_8$  of Region 1 are presented in Fig. 20 with the modulation index  $m_a = 0.3$ . When  $m_a$  becomes 0.8, the corresponding zoomed view of the switching intervals for the power devices  $S_2$ ,  $S_6$ ,  $S_7$  and  $S_8$  in Region 2-5 are shown in Fig. 21. The time interval of  $T_{ins}$  is set to  $3\mu s$ . Seen from Fig. 20 and 21, the switching transitions of different switches and the actual switching process in Sector I can be clearly identified, which fully complies with the proposed modulation scheme.

Furthermore, the experimental results of phase-A voltage, output switched current, and inductor currents  $I_{L-1}$  and  $I_{L-2}$  are shown in Fig. 22, while the FFT analysis of output filtered current in phase-A is shown in Fig. 23. As expected, the proposed converter outputs five-level switching currents. Moreover, the output filtered current and voltage are almost purely 50Hz sinusoidal waveforms with very low harmonics. This indicates that the proposed system can achieve a high-quality output with a lower power devices count compared to the prior-art MCSI topologies. Additionally, two inductor currents  $I_{L-1}$  and  $I_{L-2}$  are maintained at around 6A with almost the same current ripple, which validates that the proposed current balancing strategy also perform well.

In order to compare the efficiency performance, three additional prototypes including: a) conventional 3-level H6

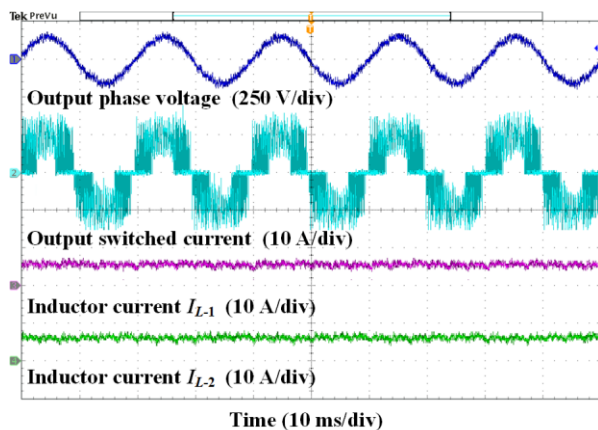


Fig. 22. Performance (experiments) of the proposed CSI topology ( $m_a$ : 0.8).

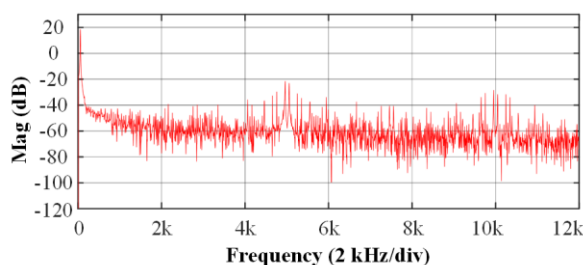


Fig. 23. FFT performance (experiments) of the proposed CSI topology (output filtered current).

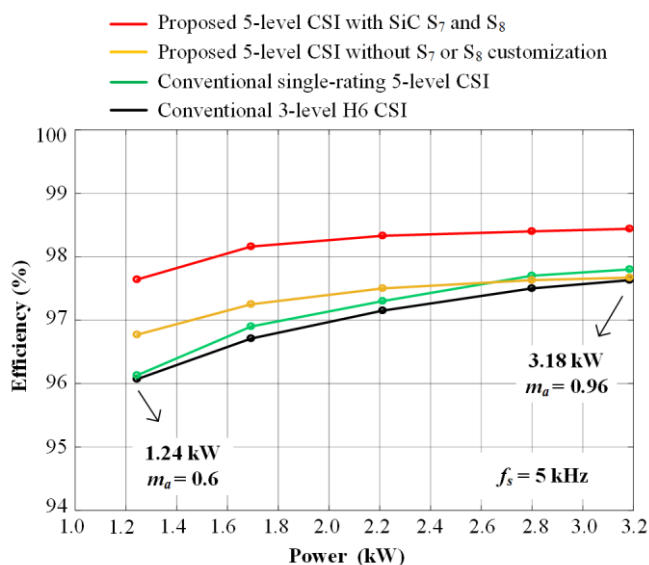


Fig. 24. Efficiency comparison of the selected converters: conventional 3-level H6 CSI, conventional single-rating inductor 5-level CSI, the proposed 5-level CSI with SiC  $S_7$  and  $S_8$ , and the proposed 5-level CSI without customization.

CSI, b) conventional single-rating inductor 5-level CSI, and c) the proposed CSI without  $S_7$  or  $S_8$  customization (i.e., silicon devices), have been built up with the same experimental parameters. More specifically, except for the SiC devices of  $S_7$  and  $S_8$  in the proposed cost-effective CSI solution, all the power switches and the power diodes are employed with the same devices (IKW20N60T and C2D20120D). Beyond that, to make a proper comparison, these converters should be evaluated under the same conditions (operating with the same output value and the same modulation index). Subsequently, a series of efficiency data of the selected converters can be recorded and compared as shown in Fig. 24. The tested power

rating is from 1.24 kW to 3.18 kW, and the modulation index  $m_a$  ranges from 0.6 to 0.96.

It can be concluded from Fig. 24 that the proposed eight-switch five-level CSI, which employs SiC MOSFETs for  $S_7$  and  $S_8$ , achieves the highest efficiency compared to the rest. At the power of 3.18 kW and the basic switching frequency ( $S_1$ - $S_6$ ) of 5 kHz, the efficiency of the proposed CSI reaches 98.46% and is almost 1% higher than that of the conventional H6 CSI. On the other hand, when the converter is configured with silicon devices for  $S_7$  or  $S_8$ , the efficiency drops as expected. In some cases, when the power rating is higher than 2.5 kW ( $m_a=0.85$ ), the efficiency is even lower than the conventional single-rating inductor 5-level CSI, as shown in Fig. 24. Despite the absence of SiC power switches, this efficiency-cross phenomenon is affected by the corresponding modulation index  $m_a$ . With the proposed modulation scheme, the switching losses of the proposed CSI may be lower than that of the conventional 5-level CSI. However, the dwell time of large vectors (conducting with  $I_{dc}$ ) will become longer with the increase of  $m_a$ , and thus leading to the increase of the conduction losses. As also shown in Fig. 24, the efficiencies of the conventional 3-level H6 CSI, conventional single-rating inductor 5-level CSI, and the proposed CSI without customized power devices for  $S_7$  or  $S_8$  are quite close. However, as demonstrated, only with SiC power devices of  $S_7$  and  $S_8$ , the proposed CSI can achieve a significant efficiency improvement. In all, the above simulations and experimental tests have validated the effectiveness of the proposed eight-switch five-level CSI topology in terms of low device count and high power quality. It thus can be a promising solution for CSI applications.

## V. CONCLUSION

This paper proposes an eight-switch three-phase five-level CSI topology. Moreover, the SVM strategy as well as its operational principle was presented in detail. In addition, the operational features and superior advantages of the proposed CSI have been discussed and benchmarked. Focusing on the unique advantages of the newly proposed topology, it performs five-level output currents, while employs with a comparable hardware cost as the three-level H6 CSI. The low output THD may help to reduce the sizes of the passive components in the system, especially the output filters. And the “small vector” can reduce the switching and conduction losses of the semiconductor switching devices in H6 CSI module, which make it possible to increase the output current rating of the system in a certain degree. The corresponding performance has been validated through simulation and experimental results.

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